Application Serial No. 09/804,830   **PATENT**
Reply to Office Action of December 28, 2004   Docket: CU-2480

APPENDIX OF ATTACHMENTS

Application Serial No. 09/804,830
Reply to Office Action of December 28, 2005

EXHIBIT I

Introduction to Balanced Receivers

A total of **FOUR** Sheets of Paper.

EXHIBIT II

The Inventor Chien CHOU's Comments On the Comparison Between the Presently
Claimed Invention and the Cited U.S. Patent No. 5,295,013 (Ono) Reference.

A total **FOUR** of Sheets of Paper.

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EXHIBIT I

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A receiver using the transformer tuning or a T-equivalent circuit of the transformer
is also effective to achieve wideband and low-noise characteristics at the same time [8].

A broadband receiver having a bandwidth from 100 MHz to 14 GHz was achieved
using a high-impedance configuration [9]. A high-performance coherent receiver with an
equivalent input noise current density of 16 pA/Hz$^{1/2}$ over a bandwidth of 6 to 15 GHz
was realized by using a T-equivalent transformer-tuned circuit configuration [10].

5.5 BALANCED RECEIVER

In coherent lightwave communication systems, high-receiver-sensitivity characteristics
can be achieved due to the shot-noise-limited operation of coherent receivers at high local
oscillator power. Generally speaking, the influence of the intensity fluctuation of signal
light on the sensitivity characteristics of coherent receivers is not so serious because the
power of the signal light is quite low in the receivers. However, the influence of the
intensity noise due to the local oscillator light cannot be disregarded due to its high power.

A new type of coherent receiver, which is referred to as a balanced receiver or dual-
detector receiver, was proposed in 1983 [11,12]. A balanced receiver suppresses the local
oscillator intensity noise quite effectively.

Figure 5.4 shows a basic configuration of a balanced receiver. The receiver consists
of one beam splitter, two photodetectors, and one differential combiner. The signal and
local oscillator light are fed into each input port of the beam splitter.

First let us assume the electrical field of the signal light, $E_s(t)$, and of the local
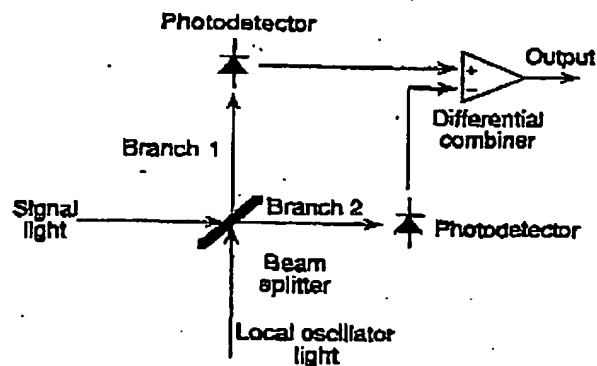oscillator light, $E_L(t)$, are

Figure 5.4 Basic configuration of a balanced receiver. (After [11]. Courtesy of The Optical Society of America.
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$$E_S(t) = \sqrt{2P_S(t)} \cos(\omega_S t + \phi_S) \qquad (5.5)$$

$$E_L(t) = \sqrt{2P_L(t)} \cos(\omega_L t + \phi_L) \qquad (5.6)$$

where $P_S(t)$, ω_S, and ϕ_S are the power, the optical angular frequency, and the optical phase of the signal light, respectively, and $P_L(t)$, ω_L, and ϕ_L are the corresponding parameters for the local oscillator light, respectively. In these assumptions, the signal and local oscillator power are functions of time, t, in which the intensity fluctuation of the signal and local oscillator power is implied.

Suppose the beam splitter has a perfect splitting ratio of 50% and the lengths of optical and electrical paths between the output of the beam splitter and the input of the differential combiner are the same for branches 1 and 2. The signal electrical field in branch 1, which is reflected by the beam splitter, is phase-shifted by $-\pi/2$ relative to the signal field that passed through the beam splitter into branch 2. The same situation applies to the local oscillator light. Hence, the electrical field of the signal and local oscillator light in each branch can be expressed as [13]

$$E_{S1}(t) = \sqrt{P_S(t)} \cos\left(\omega_S t + \phi_S - \frac{\pi}{2}\right) \qquad (5.7)$$

$$E_{S2}(t) = \sqrt{P_S(t)} \cos(\omega_S t + \phi_S) \qquad (5.8)$$

$$E_{L1}(t) = \sqrt{P_L(t)} \cos(\omega_L t + \phi_L) \qquad (5.9)$$

$$E_{L2}(t) = \sqrt{P_L(t)} \cos\left(\omega_L t + \phi_L - \frac{\pi}{2}\right) \qquad (5.10)$$

where a constant phase-shift due to transmission in the beam splitter is included in ϕ_S and ϕ_L, and a subscript for each electric field denotes the number of the branch. From (5.7) through (5.10), the generated current at the output of each PD can be calculated as

$$i_{n1} = \frac{R}{2}\{P_S(t) + P_L(t) + 2\sqrt{P_S(t)P_L(t)}[\sin((\omega_S - \omega_L)t + \phi_S - \phi_L)]\} \qquad (5.11)$$

$$i_{n2} = \frac{R}{2}\{P_S(t) + P_L(t) - 2\sqrt{P_S(t)P_L(t)}[\sin((\omega_S - \omega_L)t + \phi_S - \phi_L)]\} \qquad (5.12)$$

where R is the responsivity of each PD, which is assumed to be the same for branch 1 and for branch 2.

From (5.11) and (5.12), the output of the differential combiner becomes

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$$i_{IF} = 2R\sqrt{P_S(t)P_L(t)}[\sin(\omega_s - \omega_t)t + \phi_S - \phi_L] \qquad (5.13)$$

From (5.13), we can see that the intensity fluctuation due to the signal and local oscillator light can be effectively eliminated owing to the differential operation of the balanced receiver, because the intensity noise of the signal and local oscillator light in each branch is perfectly correlated. Some readers might wonder if in a balanced receiver the shot-noise components might also be suppressed; this is not the case. The reason is that the statistical distributions of the shot noise for each detector are independent processes; thus, the noise components can never be suppressed, even after differential combining.

There are two major advantages to using a balanced receiver, namely, the suppression of the intensity noise of the signal and local oscillator light, as described above, and the effective use of the signal and local oscillator light power in the receiver. In a receiver with one PD, half the signal and local oscillator light power are lost because we use only one branch of the receiver in Figure 5.4. On the other hand, in a balanced receiver, we can also make use of the power that has been cast in a single-detector configuration. By using a balanced receiver, we can generally achieve an improvement in receiver sensitivity of about 6 to 7 dB, compared with the single-detector configuration [10].

The preceding discussion has been given with the assumptions that the power splitting ratio of the beam splitter is exactly 50% and that the path lengths of branches 1 and 2, including optical paths and electrical paths, are completely identical. However, in practice it is difficult to match the characteristics of two PDs, and also not so easy to make the lengths of the two paths equal. Hence, the suppression capability of the common-mode noise in a balanced receiver is limited to some value. A CMRR of a balanced receiver is usually used to show the capability. The CMRR is defined as the improvement of the RIN due to the local oscillator light by the use of the balanced receiver configuration, as shown in (3.23).

Figure 5.5 shows the receiver-sensitivity improvement versus the local oscillator power using a CMRR of a balanced receiver as a parameter in a 2.5-Gb/s CPFSK heterodyne detection system. From Figure 5.5, it is seen that a CMRR of more than 30 dB is necessary to achieve a nearly shot-noise-limited operation. Figure 5.6(a) shows the calculation results of CMRR as a function of the system data rate using the delay time difference, $\Delta\tau$, between two branches as a parameter [14]. In the calculation, it is assumed that the responsivity of each PD is perfectly matched. From the results, it can be concluded that it is necessary to set $\Delta\tau$ at 1 ps for a CMRR of 30 dB at a system data rate of 2 Gb/s, which corresponds to a path difference of about 0.2 mm for a silica fiber.

Figure 5.6(b) shows the calculation results when the gain difference between two branches is taken as a parameter [14]. In the calculation, it is assumed that the delay difference between two branches is 5 ps. From the results, we can say that in low-data-rate systems, the CMRR is fixed at some value that is determined by the gain difference between two branches and that in high-data-rate systems, the CMRR is limited not by the gain difference but by the delay time difference between two branches.

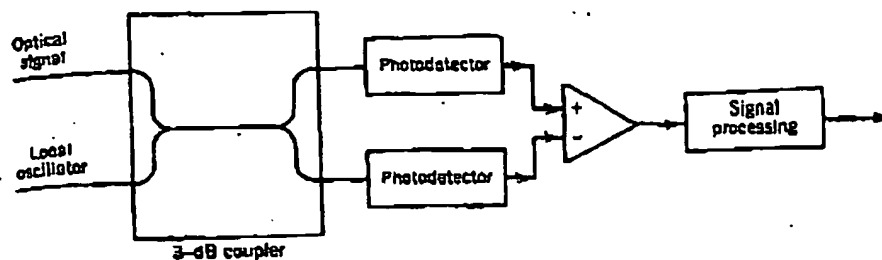EXHIBIT I   Page 3 of 4

FIGURE 6.12 A two-port balanced coherent receiver.

oscillator and splits the combined optical signal into two equal parts with an appropriate relative phase shift. The operation of a balanced receiver can be understood by considering the photocurrents I_+ and I_- generated in each branch. If the interference term in Eq. (6.1.5) or Eq. (6.1.7) has opposite signs for the two branches, the currents I_+ and I_- are given by

$$I_+ = \frac{R}{2}(P_s + P_{LO}) + R\sqrt{P_s P_{LO}}\cos(\omega_{IF}t + \phi_{IF}), \qquad (6.5.5)$$

$$I_- = \frac{R}{2}(P_s + P_{LO}) - R\sqrt{P_s P_{LO}}\cos(\omega_{IF}t + \phi_{IF}). \qquad (6.5.6)$$

where ϕ_{IF} is related to the phase difference $\phi_s - \phi_{LO}$. The subtraction of the two currents provides the heterodyne signal. The dc term is completely eliminated during the subtraction process when the two branches are balanced in such a way that each branch receives equal signal and local-oscillator powers. This occurs for a perfect 3-dB coupler with a 50% splitting ratio for each branch. The important point is that the intensity noise associated with the dc term is also eliminated during the subtraction process. The reason is that the same local oscillator provides power to each branch so that intensity fluctuations in the two branches are perfectly correlated and cancel out during subtraction of the photocurrents I_+ and I_-. It should be noted that intensity fluctuations associated with the ac term are not canceled even in a balanced receiver. However, their impact is less severe on the system performance because of the square-root dependence of the ac term on the local-oscillator power.

Balanced receivers are commonly used in the design of coherent lightwave systems because of the two advantages offered by them. First, the intensity-noise problem is nearly eliminated. Second, all of the signal and local-oscillator powers are used effectively. A single-port receiver, such as shown in Fig. 6.1, rejects parts of both P_s and P_{LO} during the mixing process. Any loss in P_s is equivalent to a power penalty. Balanced receivers use all of the signal power and avoid this power penalty. At the same time, all of the

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EXHIBIT II:

INVENTOR'S COMMENTS ON THE DIFFERENCES BETWEEN THE PRESENTLY CLAIMED INVENTION AND THE CITED U.S. PATENT NO. 5,295,013 (ONO) REFERENCE.

Regarding Claim 1 of this present application (U.S. Serial No. 09/804, 830), the differences

between this patent and U.S. Patent No. 5,295,013 (Ono) are as stated as follows:

The configuration of Ono's system is a balanced detection scheme in which two orthogonal

polarizations from PM fiber, P wave and S wave are at 45° to the geometrical axes (X-Y axes) of

PBS (see Fig. A); hence, the output signal from photo-detectors, (24) and (25) in Fig. 1 of Patent

5,295,013, becomes

$$I_x = \frac{1}{2}(A_p^2 + A_s^2) - A_p A_s \cos(\phi_p(t) - \phi_s(t)) \qquad \text{.........................(A-1)}$$
$$= (DC)_x - A_p A_s \cos(\phi_p(t) - \phi_s(t))$$

$$I_y = \frac{1}{2}(A_p^2 + A_s^2) + A_p A_s \cos(\phi_p(t) - \phi_s(t)) \qquad \text{.........................(A-2)}$$
$$= (DC)_y + A_p A_s \cos(\phi_p(t) - \phi_s(t))$$

Then, the output from differential amplifier (PA) becomes

$$\Delta I = I_y - I_x = 2A_p A_s \cos(\phi_p(t) - \phi_s(t)) \text{...(A-3)}$$

The DC terms, $(DC)_x = (DC)_y$, in Equations (A-1) and (A-2) are offset by DA automatically.

A_p and A_s are amplitudes of P and S waves respectively; $\phi_p(t)$ and $\phi_s(t)$ are phase

responses of P and S waves.   This result implies that the FSK input signal is demodulated

properly as shown in Fig. 3 of Ono's system (U.S. Patent No. 5,295,013).   However, if the

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stress or temperature variation induces PM fiber with different responses of P and S waves in

Fig. 1 of <u>Ono</u>, then the DC terms of Equations (A-1) and (A-2) becomes $(DC)_x \neq (DC)_y$. This

results in the output signal from differential amplifier:

$$\Delta I = I_x - I_y = \Delta(DC) - 2A_p A_s \cos(\phi_p(t) - \phi_s(t)) \dots\dots\dots\dots\dots\dots\dots\dots(A-4)$$

The DC terms in Equations (A-1) and (A-2) cannot be offset in Equation (A-4) where

$\Delta(DC) = (DC)_x - (DC)_y$. Then the result implies that the FSK input signal is demodulated

improperly as shown in Fig. 4 of <u>Ono</u>'s system (U.S. Patent No. 5,295,013).
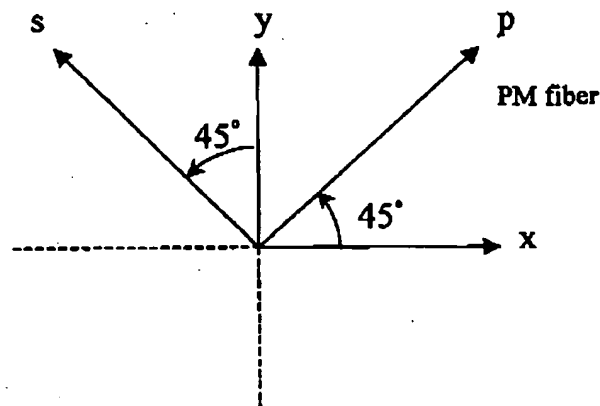

Figure A   X-Y coordinator of PBS (polarized beam splitter)

P and S are two orthogonal polarization states or two intrinsic optical axes of PM

fiber.

Therefore, in <u>Ono</u>'s system, two output signals from photo-detectors in Fig. 3 are always 180°

out of phase whatever the condition of PM fiber is.   In contrast, in our system, the

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phase-difference, $\Delta\psi = \psi_s - \psi_r$, of two output signals from photo-detectors are variable while

the equal amplitude of both heterodyne signals is required for phase-difference decoding process
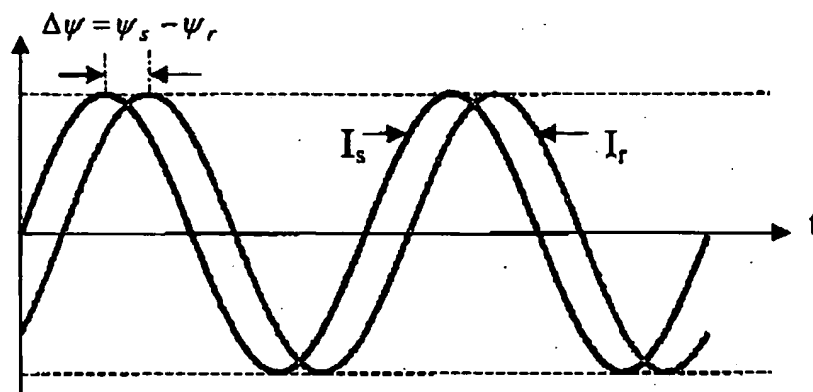
as shown in Fig. B.

Figure B   US serial # 09/804830

In summary, our system is a phase-difference decoder based on balanced detector scheme,

which uses differential amplifier (DA) in terms of the amplitude of output signal from DA.

The phase-difference between two signals from detectors is variable with time.   In our system,

the balanced detector scheme plays the role of a phase modulation (PM) to amplitude

modulation (AM) converter.   In the meantime, the DC term cancellation is resulted as well.

By contrast, Ono's system presents a technique that two output signals from detectors are

always 180° out of phase.   Besides, the phase terms in Equations (A-1) and (A-2) from

photo-detectors, (23) and (24), of Fig. 1 in Ono (U.S. Patent No. 5,295,013) are always equal.

The two intrinsic optical axes of PM fiber are required to be 45° to the axes of PBS in Fig. 1 of

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Ono's configuration (see Figure A).   Balanced detection scheme in Ono's system reduces the

DC terms in order to decode the FSK signal properly, but cannot play the role of a PM to AM

converter at the same time as our system does.   Hence, we may conclude that these two

methods are totally different not only in the system configuration and the working principle, but

also the purpose of phase decoding process.

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